PCT

(51) International Patent Classification 7 : H04L 1/00, 1/06	A1	(11) International Publication Number: **WO 00/14921**
		(43) International Publication Date: 16 March 2000 (16.03.00)

(21) International Application Number: PCT/US99/19117

(22) International Filing Date: 23 August 1999 (23.08.99)

(30) Priority Data:
60/099,212   4 September 1998 (04.09.98)   US
09/300,494   28 April 1999 (28.04.99)   US

(71) Applicant: AT & T CORP. [–/US]; 32 Avenue of the Americas, New York, NY 10013–2412 (US).

(72) Inventors: CALDERBANK, Arthur, Robert; 85 Worths Mill Lane, Princeton, NJ 08540 (US). NAGUIB, Ayman, F.; 1701 Springfield Avenue, New Providence, NJ 07974 (US). SESHADRI, Nambirajan; 88 Van Houton Avenue, Chatham, NJ 07928 (US).

(74) Agents: CONOVER, Michele, L. et al.; AT & T Corporation, P.O. Box 4110, Middletown, NJ 07748–4801 (US).

(81) Designated States: BR, CA, CN, JP, KR, MX, European patent (AT, BE, CH, CY, DE, DK, ES, FI, FR, GB, GR, IE, IT, LU, MC, NL, PT, SE).

Published
With international search report.

(54) Title: COMBINED CHANNEL CODING AND SPACE–BLOCK CODING IN A MULTI–ANTENNA ARRANGEMENT

(57) Abstract

Enhanced performance is achieved by combining channel coding with the space–time coding principles. With K synchronized terminal units transmitting on N antennas to a base station having $M \, \$m(G \, K)$ receive antennas, increased system capacity and improved performance are attained by using a concatenated coding scheme where the inner code is a space–time block code and the outer code is a conventional channel error correcting code. Information symbols are first encoded using a conventional channel code, and the resulting signals are encoded using a space–time block code. At the receiver, the inner space–time block code is used to suppress interference from the other co–channel terminals and soft decisions are made about the transmitted symbols. The channel decoding that follows makes the hard decisions about the transmitted symbols. Increased data rate is achieved by, effectively, splitting the incoming data rate into multiple channels, and each channel is transmitted over its own terminal.

1

Combined Channel Coding And Space-Block Coding
In A Multi-Antenna Arrangement

5   **Background of the Invention**

This invention relates to wireless communication and, more
particularly, to techniques for effective wireless communication in the
presence of fading, co-channel interference, and other degradations.

Physical limitations of wireless channels present a fundamental

10   technical challenge to reliable communication. Bandwidth limitations,
propagation loss, time variance, noise, interference, and multipath
fading make the wireless channel a narrow "pipe" that does not easily
accommodate the flow of data. Further challenges come from power
limitation, size, and speed of devices that are employed within portable

15   wireless apparatus.

Employing multiple transmit antennas at both the base and the
remote stations increases the capacity of wireless channels, and
information theory provides a measure of this increase. The standard
approach to exploit this capacity is linear processing at the receiver, as

20   described, for example, by J. Winters, J. Salz and R. D. Gitlin, in "The
impact of antenna diversity an the capacity of wireless communication
systems" *IEEE Trans. Communications*, Vol. 42. No. 2/3/4, pp. 1740-
1751, Feb/March/April 1994. Transmit diversity has been explored by
Wittneben in "Base station modulation diversity for digital

25   SIMULCAST," *Proc. IEEE'VTC*, pp. 505-511, May 1993, and by
Seshadri and Winters in "Two signaling schemes for improving the
error performance of frequency-division-duplex (FDD) transmission
systems using transmitter antenna diversity," *International Journal of
Wireless Information Networks*, Vol. 1, No. 1, 1994. The Wittneben

2

and Seshadri et al papers approach transmit diversity from a signal processing point of view.

Space-time codes combine signal processing at the receiver with coding techniques appropriate to multiple transmit antennas. See, for example, V. Tarokh, N. Seshadri, and A. R. Calderbank in "Space-Time Codes For High Data Rate Wireless Communication: Performance Analysis and Code Construction," *IEEE Trans. Info. Theory,* Vol. 44, No. 2, pp. 744-765, March 1998. The space-time approach provides significant gain over the aforementioned prior art. Specific space-time codes designed for 2-4 transmitting antennas perform well in slowly varying fading environments (such as indoor transmission) and come with in 2-3 dB of the theoretical outage capacity. Outage capacity is described, for example, by J. Foschini, Jr. and M. J. Gans, "On limits of wireless communication in a fading environment, when using multiple antennas," *Wireless Personal Communication,* Vol. 6, No. 3, pp. 311-335, March 1998. The bandwidth efficiency of the codes described in the Tarokh et al paper is about 3-4 times that of the current systems. The most important contributor to the improved performance is diversity, which may be thought of as providing a number of replicas of the transmitted signal to the receiver, with some replicas being less attenuated by fading. The space-time codes presented in the Tarokh et al paper provide an optimized trade-off between constellation size, data rate, diversity gain, and trellis complexity

When the number of transmit-antennas is fixed, decoding complexity (measured, for example, by the number of trellis states in the decoder) increases exponentially with transmission rate. This can be refined to some extent by designing space-time codes with a multi-level structure and adopting multi-stage decoding, as described by

3

Tarokh et al. For a moderate number of transmit antennas (3-6), this method provides higher data rates while reducing the decoding complexity. However, there is a penalty to be paid for simplified decoding. Multi-stage decoding is sub-optimal, due in part to

5   magnification in the error coefficient, and this performance penalty means that alternative solutions are needed for achieving very high data rates.

In order to achieve very high data rates on narrowband wireless channels, many antennas are needed at both the transmitter and the

10   receiver. Consider a wireless communication system employing n transmit and m receive antennas, where the subchannel between each transmit and receive antenna is quasi-static Rayleigh, flat, and independent of others. If n is fixed, then the capacity increases logarithmically with m. On the other hand, if m is fixed then intuition

15   suggests that there must come a point where adding more transmit antennas will not make much difference. Indeed, this can be seen in the mathematics of outage capacity, shown by Foschini and Gans in the aforementioned paper. Thus, it turns out that in the presence of one receive antenna little can be gained in terms of outage capacity by using

20   more than 4 transmit antennas. A similar argument shows that if there are two receive antennas, using 6 transmitting antennas provides almost all of the capacity increase that can be obtained.

If n increases and $m \geq n$, then information theory shows that the capacity of the system increases at least linearly as a function of n.

25   Thus, it makes sense to increase the number of antennas at both receiver and transmitter in order to obtain higher capacities. Employing many antennas at both transmitter and receiver creates a multiple-input multiple-output system where the number of degrees of freedom is given by the product of the number of transmit and receive antennas.

4

Foschini considered such a system in "Layered space-time architecture for wireless communication in a fading environment when using multi-element antennas," *Bell Labs Technical Journal,* Vol. 1, No. 2, Autumn 1996. He proposed a multi-layered structure that, in

5   principle, can achieve a tight lower bound on the capacity. If n transmit and n receive antennas are used, then at the receiver the transmitted signal from transmit antenna 1 is treated as the desired signal, while the signals transmitted from other transmit antennas are treated as interference. Linear processing is then used to suppress the interfering

10   signals using n receive-antennas, providing a diversity gain of one. Once the signal transmitted from antenna 1 is correctly detected, the signal transmitted from antenna 2 is treated as the desired signal while the signals transmitted from transmit antennas 3, 4, ..., n are treated as interference. The contribution of the signal transmitted from antenna 1,

15   now that it has been detected, is subtracted from the signal received at receiver antennas 1 through n. Thereafter, detection of the signal transmitted by antenna 2 proceeds with linear processing that is applied to suppress the interfering signals from antennas 3 through n. This provides a diversity gain of two. This process is repeated until all the

20   transmitted signals are detected. Clearly, the worst case diversity in this architecture is one. For such a system, long frames of data combined with powerful coding techniques are needed to achieve the lower bound on outage capacity.

In U.S.A. Application Serial No. 09/114838 filed July 14, 1998,

25   claiming priority of U.S. A. Provisional Application Number 60/052689 filed July 16, 1997, an arrangement is disclosed that provides enhanced performance which is realized by employing a perspective that combines array signal processing with channel coding. Specifically, antennas at the transmitter are partitioned into small

5

groups, and individual space-time codes are used to transmit

information from each group of antennas. At the receiver, an individual

space-time code is decoded by a linear array processing technique that

suppresses signals transmitted by other groups of antennas by treating

5   them as interference. The contribution of the decoded signal to other

received signal is then subtracted from those received signals. What

results is a simple receiver structure that provides diversity and coding

gain over uncoded systems with a given diversity gain. This

combination of array processing at the receiver and coding techniques

10   for multiple transmit antennas provides reliable and very high data rate

communication over wireless channels. One advantage of the group

interference suppression method over the Foschini architecture is that

the number of receive antennas can be less than the number of transmit

antennas.

15   In U.S.A. Application Serial No. 09/149,163 filed Sep. 4, 1999,

claiming priority of U.S. A. Provisional Application Number

60/052689 filed July 17, 1997, an arrangement is disclosed wherein K

synchronized terminal units transmit on N antennas to a base station

having $M \geq K$ antennas. An improvement is achieved by employing

20   both interference cancellation (IC) and maximum likelihood (ML)

decoding. More specifically, space-time block coding is employed in

transmitters that employ N transmit antennas each, and the signals are

received in a receiver that employs M receiving antennas. By

exploiting the structure of the space-time block code, K-1 interfering

25   transmitting units are cancelled at the receiver, regardless of the number

of transmitting antennas, N, when decoding the signals transmitted by a

given mobile unit. Also disclosed is an arrangement where signals of a

first terminal unit are decoded first, and the resulting decoded signals

are employed to cancel their contribution to the signals received at the

6

base station antennas while decoding the signals of the remaining K-1

terminal units. The process is repeated among the remaining K-1

terminal units.

5   **Summary of the Invention**

Enhanced performance is achieved by combining channel

coding with the space-time coding principles disclosed in the '163

applications. More specifically, with K *synchronized* terminal units

transmitting on N antennas to a base station having $M \geq K$ receive

10   antennas, increased system capacity and improved performance are

attained by using a concatenated coding scheme where the inner code is

a space-time block code and the outer code is a conventional channel

error correcting code. That is, information symbols are first encoded

using a conventional channel code. The channel code is then encoded

15   signal is then encoded using a space-time block code, and transmitted

over N antennas. At the receiver, the inner space-time block code is

used to suppress interference from the other co-channel terminals and

soft decisions are made about the transmitted symbols. The channel

decoding that follows makes the hard decisions about the transmitted

20   symbols.

Increased data rate is achieved by, effectively, splitting the

incoming data rate into multiple channels, and each channel is

transmitted over its own terminal. Viewed another way, information

symbols from a transmitting terminal is split into L parallel streams.

25   Stream l is encoded using a channel code with rate R_l and then coded

with a space-time block encoder with N transmitting antennas.

Advantageously, the coding rates are chosen such that

$R_1 > R_2 > \cdots > R_L$.

7

Brief Description of the Drawing

FIG. 1 depicts an arrangement that, illustratively, includes a base station 30 with four antennas, a terminal units 20 with two antennas, and a terminal unit 20 with two antennas; and

5   FIG. 2 illustrates a terminal unit that splits the input signal into two streams, and each stream is transmitted over a separate two-antenna arrangement.

Detailed Description

10   FIG. 1 illustrates a unit 10 that employs a space-time block coding unit 13 that is followed by a conventional constellation mapper and pulse shaping circuit 16. The output of circuit 16 is fed to two transmitting antennas 11 and 12. The input symbols to the space-time block encoder are divided into groups of two symbols each, and at a

15   given symbol period, the two symbols in each group $\{c_1, c_2\}$ are transmitted simultaneously from the two antennas. The signal transmitted from antenna 11 is c_1 and the signal transmitted from antenna 12 is c_2. In the next symbol period, the signal $-c_2^*$ is transmitted from antenna 11 and the signal c_1^* is transmitted from

20   antenna 12.

In receiver 20, signals are received by antennas 21 and 22 and are applied to detector 25. Channel estimators 23 and 24 operate on the incoming signal of antennas 21 and 24, respectively, in a conventional manner to develop estimates of the channel parameters. Those

25   estimates are applied to detector 25. In the mathematical development of the algorithms disclosed herein, it is assumed that the channel from each of the two transmit antennas remains fixed over a two consecutive symbol periods. That is,

8

$$h_i(nT) = h_i((n+1)T), \quad i = 1,2. \tag{1}$$

To ascertain the channel characteristics, the transmitter carries out a calibration session, during which pilot signals or tones are transmitted. It is the signals that are received during the calibration session that are

5   utilized by the channel estimator circuits 23 and 24, which are well known circuits.

Maximum Likelihood Detection

The received signals at antenna 21 can be expressed as

$$r_1 = h_1 c_1 + h_2 c_2 + \eta_1 \tag{2}$$

10

$$r_2 = -h_1 c_2^* + h_2 c_1^* + \eta_2, \tag{3}$$

where r_1 and r_2 are the received signals over two consecutive symbol periods, h_1 denotes the fading channel between transmit antenna 11 and receive antenna 21, h_2 denotes channel between transmit antenna 12 and receive antenna 21, and η_1 and η_2 are noise terms, which are

15   assumed to be complex Gaussian random variables with zero mean and power spectral density $N_0/2$ per dimension. Defining the vectors

$\mathbf{r} = [r_1 r_2^*]^T$, $\mathbf{c} = [c_1 c_2]^T$, and $\eta = [\eta_1 \eta_2^*]^T$, equations (2) and (3) can be rewritten in a matrix form as

$$\mathbf{r} = \mathbf{H} \cdot \mathbf{c} + \eta, \tag{4}$$

20   where the channel matrix $\mathbf{H}$ is defined as

$$\mathbf{H} = \begin{bmatrix} h_1 & h_2 \\ h_2{}^* & -h_1{}^* \end{bmatrix}. \tag{5}$$

The vector η is a complex Gaussian random vector with zero mean and covariance $N_0 \cdot \mathbf{I}$. Defining C as the set of all possible symbol pairs $\mathbf{c} = \{c_1, c_2\}$, and assuming that all symbol pairs are equi-probable, it

25   can be easily shown that the optimum maximum likelihood (ML) decoder selects from C the symbol pair $\hat{\mathbf{c}}$ that minimizes the expression

9

$\|\mathbf{r} - \mathbf{H} \cdot \hat{\mathbf{c}}\|^2$. This can be written as

$$\hat{\mathbf{c}} = \arg \min_{\hat{\mathbf{c}} \in C} \|\mathbf{r} - \mathbf{H} \cdot \hat{\mathbf{c}}\|^2 . \qquad (6)$$

It was shown by S. Alamouti in "Space Block Coding: A simple Transmitter Diversity Scheme for wireless Communications,"

5   submitted to IEEE JSAC, September 1997, that the diversity order of the above space-time block code is equivalent to that of a two branch maximal ratio receive combining (MRRC). Alamouti also showed that, because of the orthogonality of the matrix $\mathbf{H}$, this decoding rule decomposed into two separate decoding rules for c_1 and c_2. The

10   *uncertainty*, Δ_c, of the decoded symbols $\hat{\mathbf{c}}$ is defined as

$$\Delta_c = \|\mathbf{r} - \mathbf{H} \cdot \hat{\mathbf{c}}\|^2 . \qquad (7)$$

The maximum likelihood (ML) rule of equation (6) can be simplified by realizing that the channel matrix $\mathbf{H}$ is orthogonal; that is, that $\mathbf{H} * \mathbf{H} = (|h_1|^2 + |h_2|^2)\mathbf{I}$. This yields a modified received vector,

15

$$\tilde{\mathbf{r}} = \mathbf{H} * \mathbf{r} = (|h_1|^2 + |h_2|^2) \cdot \mathbf{c} + \tilde{\eta}, \qquad (8)$$

where $\tilde{\eta} = \mathbf{H} * \eta$. This leads to

$$\hat{\mathbf{c}} = \arg \min_{\hat{\mathbf{c}} \in C} \|\tilde{\mathbf{r}} - (|h_1|^2 + |h_2|^2) \cdot \hat{\mathbf{c}}\|^2 . \qquad (9)$$

Thus it follows that by using a simple linear combining, the decoding rule of equation (9) reduces to two separate, and much simpler,

20   decoding rules for c_1 and c_2. When using a signaling constellation with 2b constellation points, this reduces the number of decoding metrics that have to be computer for ML decoding from 2^{2b} to 2×2^b.

When receiver 20 uses M receiver antennas, the received vector at antenna m is

25

$$\mathbf{r}_m = \mathbf{H}_m \cdot \mathbf{c} + \eta_m , \qquad (10)$$

10

where the channel matrix $\mathbf{H}_m$ is defined as

$$\mathbf{H}_m = \begin{bmatrix} h_{1m} & h_{2m} \\ h_{2m}^* & -h_{1m}^* \end{bmatrix}.$$

(11)

In this case, the optimum ML decoding rule is

$$\hat{\mathbf{c}} = \arg \min_{\hat{\mathbf{c}} \in C} \sum_{m=1}^{M} \left\| \mathbf{r}_m - \mathbf{H}_m \cdot \hat{\mathbf{c}} \right\|^2,$$

(12)

5   and the corresponding *uncertainty*, Δ_c, of the decoded symbols $\hat{\mathbf{c}}$ is

defined by

$$\Delta_c = \sum_{m=1}^{M} \left\| \mathbf{r}_m - \mathbf{H}_m \cdot \hat{\mathbf{c}} \right\|^2.$$

(13)

As before, in the case of M receive antennas, the decoding rule can be

simplified by premultiplying the received signal by $\mathbf{H}_m^*$.

10   As indicated above, FIG. 1 shows two terminal units 10 and 30,

and the issue that needs to be addressed is the detection performance at

the base station receiver when the two terminal units synchronously

transmit *over the same time and frequency channel*.

In the notation below, g_{11} denotes the fading channel between

15   transmit antenna 31 and receive antenna 21, g_{12} denotes the channel

between antenna 31 and antenna 22, g_{21} denotes the channel between

antenna 32 and antenna 21, and g_{22} denotes the channel between

antenna 32 and antenna 22. Also, $\{c_1, c_2\}$ and $\{s_1, s_2\}$ denote the two

symbols transmitted from terminal units 10 and 30, respectively.

20   At receiver 20, the received signals over two consecutive

symbol periods at receive antenna 21, r_{11} and r_{12}, are

$$r_{11} = h_{11}c_1 + h_{21}c_2 + g_{11}s_1 + g_{21}s_2 + \eta_{11}$$

(14)

$$r_{12} = -h_{11}c_2^* + h_{21}c_1^* - g_{11}s_2^* + g_{21}s_1^* + \eta_{12}.$$

(15)

Defining $\mathbf{r}_1 = [r_{11} r_{12}{}^*]^T$, $\mathbf{c} = [c_1\ c_2]^T$, $\mathbf{s} = [s_1\ s_2]^T$, and $\mathbf{n}_1 = [\eta_{11}\ \eta_{12}^*]^T$

11

equations (14) and (15) can be rewritten in matrix form as

$$\mathbf{r}_1 = \mathbf{H}_1 \cdot \mathbf{c} + \mathbf{G}_1 \cdot \mathbf{s} + \mathbf{n}_1, \tag{16}$$

where the channel matrices $\mathbf{H}_1$ and $\mathbf{G}_1$ between the transmitter units

10 and 30 and receive antenna 21 are given by

5
$$\mathbf{H}_1 = \begin{bmatrix} h_{11} & h_{21} \\ h_{21}^* & -h_{11}^* \end{bmatrix}, \text{ and } \mathbf{G}_1 = \begin{bmatrix} g_{11} & g_{21} \\ g_{21}^* & -g_{11}^* \end{bmatrix}. \tag{15}$$

The vector $\mathbf{n}_1 = [\eta_{11}\eta_{12}{}^*]^T$ is a complex Gaussian random vector with

zero mean and covariance $N_0 \cdot \mathbf{I}$. Similarly, the received signals over

two consecutive symbol periods at receive antenna 22, r_{21} and r_{22}, are

$$r_{21} = h_{12}c_1 + h_{22}c_2 + g_{12}s_1 + g_{22}s_2 + \eta_{21} \tag{16}$$

10
$$r_{22} = -h_{12}c_2^* + h_{22}c_1^* - g_{12}s_2^* + g_{22}s_1^* + \eta_{22}. \tag{17}$$

In a similar fashion, defining $\mathbf{r}_2 = [r_{21}r_{22}{}^*]^T$ and $\mathbf{n}_2 = [\eta_{21}\eta_{22}{}^*]^T$

equations (16) and (17) can be rewritten as

$$\mathbf{r}_2 = \mathbf{H}_2 \cdot \mathbf{c} + \mathbf{G}_2 \cdot \mathbf{s} + \mathbf{n}_2, \tag{18}$$

where the channel matrices $\mathbf{H}_2$ and are given by

15
$$\mathbf{H}_2 = \begin{bmatrix} h_{12} & h_{22} \\ h_{22}^* & -h_{12}^* \end{bmatrix}, \text{ and } \mathbf{G}_2 = \begin{bmatrix} g_{12} & g_{22} \\ g_{22}^* & -g_{12}^* \end{bmatrix}. \tag{19}$$

Equations (14) and (18) can be combined to yield the matrix form

$$\mathbf{r} = \begin{bmatrix} \mathbf{r}_1 \\ \mathbf{r}_2 \end{bmatrix} = \begin{bmatrix} \mathbf{H}_1 & \mathbf{G}_1 \\ \mathbf{H}_2 & \mathbf{G}_2 \end{bmatrix} \begin{bmatrix} \mathbf{c} \\ \mathbf{s} \end{bmatrix} + \begin{bmatrix} \mathbf{n}_1 \\ \mathbf{n}_2 \end{bmatrix}. \tag{20}$$

Minimum Mean-Squared Error Interference Cancellation (MMSE IC)

When seeking to detect and decode signals $\{c_1, c_2\}$ by

20 minimizing a mean-squared error criterion, the goal is find a linear

combination of the received signals such that the mean-squared error in

detecting the signals $\{c_1, c_2\}$ is minimized. In general terms, this can

be expressed by an error cost function that is to be minimized, such as

the function

12

$$J(\alpha,\beta) = \left\| \sum_{i=1}^{4} \alpha_i^* r_i - (\beta_i^* c_1 + \beta_i^* c_2) \right\|^2 = \left\| \alpha^* \cdot \mathbf{r} - \beta^* \cdot \mathbf{c} \right\|^2 ,$$

(21)

where $\mathbf{r} = \begin{bmatrix} r_1 & r_2 & r_3 & r_4 \end{bmatrix}^T = \begin{bmatrix} r_{11} & r_{12} & r_{21} & r_{22} \end{bmatrix}^T$.

One may note that a minimum is certainly reached when both α and β are equal to 0, but that, of course, is not desired. Therefore, either β_1 or β_2 is set to 1.

When β_1 is set to 1, we get the following minimization criterion from equation (40)

$$J_1(\alpha_1,\beta_1) = \left\| \sum_{i=1}^{5} \alpha_{1i}^* r_{1i} - c_1 \right\|^2 = \left\| \widetilde{\alpha}_1^* \widetilde{r}_1 - c_1 \right\|^2 = J_1(\widetilde{\alpha}_1),$$

(22)

where $\widetilde{\alpha}_1 = [\alpha_{11},\alpha_{12},\alpha_{13},\alpha_{14},-\beta_2] = \begin{bmatrix} \alpha_1 & -\beta_2 \end{bmatrix}$ and $\widetilde{r}_1 = \begin{bmatrix} \mathbf{r}^T & c_2 \end{bmatrix}^T$.

From this it can be seen that

$$\widetilde{r}_1 = \begin{bmatrix} \mathbf{H} & \mathbf{0} \\ \mathbf{0}^T & 1 \end{bmatrix} \begin{bmatrix} \widetilde{c} \\ c_2 \end{bmatrix} + \begin{bmatrix} \mathbf{n} \\ 0 \end{bmatrix} = \mathbf{R} \cdot \mathbf{d}_1 + \widetilde{\eta},$$

(23)

where $\mathbf{0} = \begin{bmatrix} 0 & 0 & 0 & 0 \end{bmatrix}^T$.

What is needed is to select $\widetilde{\alpha}_1$ so that the expected value of the expression in equation (22) is minimized. That is, select $\widetilde{\alpha}_1$ to minimize

$$E\{J_1(\widetilde{\alpha}_1)\} = E\{J_1(\widetilde{\alpha}_1)\} = E\{(\widetilde{\alpha}_1^* \widetilde{r}_1 - c_1)(\widetilde{\alpha}_1^* \widetilde{r}_1 - c_1)^*\}$$
$$= \widetilde{\alpha}_1^* E\{\widetilde{r}_1 \widetilde{r}_1^*\}\widetilde{\alpha}_1 + E\{c_1 c_1^*\} - E\{\widetilde{r}_1 c_1^*\}\widetilde{\alpha}_1^* - E\{c_1 \widetilde{r}_1^*\}\widetilde{\alpha}_1$$

(24)

Taking the partial derivative with respect to $\widetilde{\alpha}_1$ and setting it to zero results in

$$\begin{bmatrix} \mathbf{M} & \mathbf{h}_2 \\ \mathbf{h}_2^* & 1 \end{bmatrix} \begin{bmatrix} \alpha_1^* \\ -\beta_2 \end{bmatrix} = \begin{bmatrix} \mathbf{h}_1 \\ 0 \end{bmatrix},$$

(25)

where $\mathbf{M} = \mathbf{HH}^* + \frac{1}{\Gamma}\mathbf{I}$, Γ is the signal to noise ratio, $\mathbf{I}$ is the 4 by 4

identity matrix, $\mathbf{h}_1$ is the first column of $\mathbf{H}$, and $\mathbf{h}_2$ is the second column

of $\mathbf{H}$. It follows that

$$\alpha_I = \left(\mathbf{M} - \mathbf{h}_2\mathbf{h}_2^*\right)^{-1}\mathbf{h}_1 \quad \text{and} \quad \beta_2^* = \mathbf{h}_2^*\left(\mathbf{M} - \mathbf{h}_2\mathbf{h}_2^*\right)^{-1}\mathbf{h}_1. \tag{26}$$

It can be shown that

$$(\mathbf{M} - \mathbf{h}_2\mathbf{h}_2^*)^{-1} = \mathbf{M}^{-1} + \frac{\mathbf{M}^{-1}\mathbf{h}_2\mathbf{h}_2^*\mathbf{M}^{-1}}{1 - \mathbf{h}_2^*\mathbf{M}^{-1}\mathbf{h}_2}, \tag{27}$$

which yields

$$\beta_2^* = \frac{\mathbf{h}_2^*\mathbf{M}^{-1}\mathbf{h}_1}{1 - \mathbf{h}_2^*\mathbf{M}^{-1}\mathbf{h}_2}. \tag{28}$$

From the structure of the matrix $\mathbf{H}$ we can easily verify that $\mathbf{h}_1$ and $\mathbf{h}_2$

are orthogonal. Using this fact and the structure of the matrix $\mathbf{M}$, it can

be shown that

$$\beta_2 = 0 \tag{29}$$

$$\alpha_1 = \mathbf{M}^{-1}\mathbf{h}_1. \tag{30}$$

Hence, the MMSE IC solution given by equations (29) and (30) will

minimize the mean-squared error in c_1 without any regard to c_2.

Considering the alternative cost function when β_2 is set to 1, a similar

analysis leads to the conclusion that

$$\beta_1 = 0 \tag{31}$$

$$\alpha_2 = \mathbf{M}^{-1}\mathbf{h}_2 \tag{32}$$

In this case, the MMSE IC solution given by equations (31) and (32)

will minimize the mean-squared error in c_2 without any regard to c_1.

Therefore, from equations (29)-(32), we can easily see that the MMSE

interference canceller for signals from terminal unit 10 will consist of

two different sets of weights α_1 and α_2 for c_1 and c_2, respectively. The

14

weights for decoding signals from terminal 30 can be obtained in a
similar fashion, as expected. Thus, the decoding of signals from
terminal units 10 and 30 can be performed with a single subroutine
MMSE.DECODE in decoder 25 as follows:

$$(\mathbf{c}, \Delta_c) = \text{MMSE.DECODE}(\mathbf{r}_1, \mathbf{r}_2, \mathbf{H}_1, \mathbf{H}_2, \mathbf{G}_1, \mathbf{G}_2, \Gamma)$$

{

$$\tilde{\mathbf{r}} = \begin{bmatrix} \mathbf{r}_1^T & \mathbf{r}_1^T \end{bmatrix}^T$$

$$\tilde{\mathbf{H}} = \begin{bmatrix} \mathbf{H}_1 & \mathbf{G}_1 \\ \mathbf{H}_2 & \mathbf{G}_2 \end{bmatrix}$$

$$\mathbf{M} = \mathbf{H}\mathbf{H}^* + \frac{1}{\Gamma}\mathbf{I}$$

$$\mathbf{h}_1 = \begin{bmatrix} \mathbf{h}_{11} & \mathbf{h}_{21}^* & \mathbf{h}_{12} & \mathbf{h}_{22}^* \end{bmatrix}^T = \text{first column of } \mathbf{H}$$

$$\mathbf{h}_2 = \begin{bmatrix} \mathbf{h}_{21} & -\mathbf{h}_{11}^* & \mathbf{h}_{22} & -\mathbf{h}_{21}^* \end{bmatrix}^T = \text{second column of } \mathbf{H}$$

$$\alpha_1^* = \mathbf{M}^{-1}\mathbf{h}_1 , \ \alpha_2^* = \mathbf{M}^{-1}\mathbf{h}_2$$

$$\hat{\mathbf{c}} = \underset{\hat{c}_1, \hat{c}_2 \in C}{\arg\min} \ \{\|\alpha_1^*\tilde{\mathbf{r}} - \hat{c}_1\|^2 + \|\alpha_1^*\tilde{\mathbf{r}} - \hat{c}_2\|\}$$

$$\Delta_c = \|\alpha_1^*\tilde{\mathbf{r}} - \hat{c}_1\|^2 + \|\alpha_1^*\tilde{\mathbf{r}} - \hat{c}_2\|^2$$

5   }

With such a subroutine, both $\hat{\mathbf{s}}$ and $\hat{\mathbf{c}}$ can be estimated, as follows:

$$(\hat{c}, \Delta) = \text{MMSE.DECODE}(r_1, r_2, H_1, H_2, G_1, G_2, \Gamma) \qquad (33)$$

$$(\hat{s}, \Delta) = \text{MMSE.DECODE}(r_1, r_2, G_1, G_2, H_1, H_2, \Gamma) \qquad (34)$$

Additional improvement can be realized by employing a two-
10   step interference cancellation approach. In this two-step approach, the
receiver decodes signals from both terminals using the subroutine
MMSE.DECODE disclosed above. Assuming that symbols from the
terminal unit 10 $\hat{c}_o$ have been decoded correctly, the receiver can, then,
perfectly cancel the contribution of the terminal unit 10 in the received
15   signal vectors $\mathbf{r}_1$ and $\mathbf{r}_2$. The receiver then uses $\mathbf{x}_1$ and $\mathbf{x}_2$, the received
signal vectors after canceling signals from terminal unit 10, to *re-*

15

decode symbols $\hat{s}_o$ from terminal unit 30 using the optimum ML

decoding rule in equation (10). Assuming that the symbols from

terminal unit 10 have been decoded correctly, the performance for

terminal unit 30 will be equivalent to that with 2 transmit and 2 receive

5   antennas (which is equivalent to 4 branch MRC diversity). The

receiver then repeats the above step assuming that symbols from

terminal unit 30 $\hat{s}_1$ have been decoded correctly using the

MMSE.DECODE subroutine. As before, the receiver cancels the

contribution of terminal unit 30 in the received signal vectors r_1 and

10   uses y_1 and y_2, the received signal vectors after cancelling signals from

terminal unit 30, *to re-decode* symbols from terminal unit 10 $\hat{c}_1$ using

the optimum ML decoding rule in equation (10). Also as before,

assuming that symbols from terminal unit 30 have been decoded

correctly, the performance for terminal unit 10 will be equivalent to that

15   with 2 transmit and 2 receive antennas. Letting $\Delta_0 = \Delta_{c_0} + \Delta_{s_0}$ and

$\Delta_1 = \Delta_{c_1} + \Delta_{s_1}$ denote the overall uncertainty for $\hat{c}_0$ and $\hat{s}_0$, and for $\hat{c}_1$

and $\hat{s}_1$, respectively, the receiver compares the overall uncertainty and

chooses the pair $(\hat{c}_o, \hat{s}_o)$ if $\Delta_o < \Delta_1$, and $(\hat{c}_1, \hat{s}_1)$ otherwise. The two-step

interference cancellation and ML decoding algorithm is presented

20   below in pseudo-code subroutine II.MMSE.DECODE.

16

$$(\hat{c}, \hat{s}) = \text{II.DECODE}(r_1, r_2, H_1, H_2, G_1, G_2, \Gamma)$$

{

$$(\hat{c}_0, \Delta_{c,o}) = \text{MMSE.DECODE}(r_1, r_2, H_1, H_2, G_1, G_2, \Gamma)$$

$$x_1 = r_1 - H_1 \cdot \hat{c}_o \quad , \quad x_2 = r_2 - H_2 \cdot \hat{c}_o$$

$$F(s) = \|x_1 - G_1 \cdot s\|^2 + \|x_2 - G_2 \cdot s\|^2$$

$$\hat{s}_o = \arg\min_{s \in S} (F(s)) \quad , \quad \Delta_{s,o} = F(s)$$

$$(\hat{s}_1, \Delta_{s,1}) = \text{MMSE.DECODE}(r_1, r_2, G_1, G_2, H_1, H_2, \Gamma)$$

$$y_1 = r_1 - G_1 \cdot \hat{s}_1 \quad , \quad y_2 = r_2 - G_2 \cdot \hat{s}_1$$

$$F(c) = \|y_1 - H_1 \cdot c\|^2 + \|y_2 - H_2 \cdot c\|^2$$

$$\hat{c}_1 = \arg\min_{c \in C} (F(c)) \quad , \quad \Delta_{c,1} = F(c)$$

$$\text{If } (\Delta_{c,o} + \Delta_{s,o}) < (\Delta_{c,1} + \Delta_{s,1})$$

$$(\hat{c}, \hat{s}) = (\hat{c}_o, \hat{s}_o)$$

Else

$$(\hat{c}, \hat{s}) = (\hat{c}_1, \hat{s}_1)$$

}

With appreciation of the above-disclosed theoretical background, we realized that enhanced performance is attained by devoting the space-time block coding to interference cancellation and

5   ML decoding, while another coding schema may be used to overcome channel-caused degradations, such as fading. Accordingly, each of the transmitters in FIG. 1 includes a channel coder (14 and 34, respectively) that is interposed between the input signal and the transmitter's space-time block coder. Channel coders 14 and 24 can

10   employ any conventional channel error correcting code (for example a trellis code, or a convolutional code).

At receiver 20, the inner space-time block code is decoded in element 26 and is used to suppress interference from the various co-channel terminals, using the MMSE approach disclosed above.

17

Element 26 forms the two interference canceling vectors α_{i1} and α_{i1}

corresponding to some terminal, i, and element 27 forms the two

decision variables

$$\xi_1 = \alpha_1^* r \quad and \quad \xi_2 = \alpha_2^* r \tag{35}$$

5   Those decisions, however, are used as soft decisions of the transmitted

information symbols, and are fed to channel decoder 28, which is a

conventional decoder that corresponds to the type of encoding

performed in channel encoders 14 and 34. Thus, in the arrangement

depicted in FIG. 1, the structure of an inner coder is used for

10   interference suppression, such that many co-channel terminals can

operate simultaneously while providing diversity. The output of the

inner code space-time decoder forms the input to the outer coder

decoder, which decides on the transmitted information while providing

protection against channel errors.

15   FIG. 2 presents an arrangement for increasing the data rate or

throughput in wireless systems. In FIG. 2, the information to be

transmitted is demultiplexed in element 40 into two streams. One

stream is applied to channel encoder 41, and the other stream is applied

to channel encoder 51. The output of channel encoder 41 is applied to

20   space-time block encoder 42, and then to mapper and pulse shaper 43

and antennas 44 and 45. Similarly, the output of channel encoder 51 is

applied to space-time block encoder 52, and then to mapper and pulse

shaper 53 and antennas 54 and 45. Generalizing, information symbols

from a transmitting terminal are split into L parallel streams. Stream l

25   is then encoded using a channel code with rate R_l and then coded with a

space-time block encoder with N transmitting antennas. The coding

rates can be the same, but an advantage accrues when the coding rates

are chosen such that $R_1 > R_2 > \cdots > R_L$. In such a case, symbols

transmitted in stream l will have better immunity against channel errors

18

as compared to symbols transmitted in stream u where $u>l$. The base station receiver is assumed to be equipped with at least L receive antennas. The base station receiver treats each stream as a different user and uses the iterative interference cancellation techniques

5   disclosed above, or the ones disclosed in the aforementioned '163 application. Since the first stream has the smallest coding rate R_l, it has the best immunity against the channel errors and most likely it will be error free. The receiver then uses the decoded symbol of stream l to subtract the contributions of the first stream in the total received

10   signals, while decoding the remaining $L-l$ streams. In decoding the remaining $L-l$ streams, the decoder decodes signals from the second stream first, since it has the best immunity against channel errors among the remaining $L-l$ streams (due to its lowest rate, R_2 from among the remaining streams). Then the receiver uses the decoded

15   symbols for the second stream to cancel out its contribution in the received signal. This process is repeated until all streams are decoded.

It can be shown that, in this case, the system throughput is given by

$$\rho = \frac{1}{L}\sum_{l=1}^{L} R_l(1 - FER_l), \qquad (36)$$

20   where FER_l is the frame error rate of stream l.

19

Claims:

1. An arrangement comprising:

a channel code encoder responsive to an applied input signal,

a space-time encoder responsive to output signal of said channel

code encoder; and

a modulator responsive to said space time-encoder.

2. The arrangement of claim 1 further comprising pulse shaping circuitry and at least two antennas for transmitting a space-time coded signal created by said space-time encoder and modulated by said modulator.

3. A transmitter comprising:

a demultiplexer responsive to an applied input signal for developing a plurality of at least two signal streams, and

a like plurality of channel coding/space-time coding transmitters, each responsive to a different signal stream of said plurality of signal streams.

4. The transmitter of claim 3 where each of said channel coding/space-time coding transmitters comprises:

a channel coder of rate R_i,

a space-time encoder responsive to output signal of said channel code encoder,

a modulator responsive to said space time-encoder,

pulse shaping circuitry responsive to said modulator, and

at least two antennas for transmitting a space-time coded signal created by said space-time encoder, modulated by said modulator, and conditioned by said pulse shaping circuitry.

5. The transmitter of claim 4 where said demultiplexer develops an L plurality of signal streams, where said channel coders in said L channel coding/space-time coding transmitters develop rates R_i

5   $i=1,2,...,L$, that are not identical to each other.

6. The transmitter of claim 4 where said demultiplexer develops an L plurality of signal streams, where said channel coders in said L channel coding/space-time coding transmitters develop rates R_i

10   $i=1,2,...,L$, that are such that $R_1 > R_2 > \cdots > R_L$.

7. The transmitter of claim 1 where said channel code encoder performs trellis encoding.

15   8. The transmitter of claim 1 where said channel code encoder performs convolutional encoding.

9. A receiver comprising:

a detector of space-time coded signal; and

20   a decoder for decoding a channel code encoded signal that is embedded in output signals of said detector.

10. The receiver of claim 9 where said detector employs a MMSE IC decoder.

25

11. The receiver of claim 9 where said detector employs a two step algorithm to develop a weights vector for canceling interfering signals from terminals other than a given terminal whose signal is being detected.

21

12. The receiver of claim 11 where said two step algorithm is:

$$(\hat{c}, \hat{s}) = \text{II.DECODE}(r_1, r_2, H_1, H_2, G_1, G_2, \Gamma)$$

{

$$(\hat{c}_0, \Delta_{c,o}) = \text{MMSE.DECODE}(r_1, r_2, H_1, H_2, G_1, G_2, \Gamma)$$

$$x_1 = r_1 - H_1 \cdot \hat{c}_o \quad , \quad x_2 = r_2 - H_2 \cdot \hat{c}_o$$

$$F(s) = \|x_1 - G_1 \cdot s\|^2 + \|x_2 - G_2 \cdot s\|^2$$

$$\hat{s}_o = \underset{s \in S}{\arg\min} \, (F(s)) \quad , \quad \Delta_{s,o} = F(s)$$

$$(\hat{s}_1, \Delta_{s,1}) = \text{MMSE.DECODE}(r_1, r_2, G_1, G_2, H_1, H_2, \Gamma)$$

$$y_1 = r_1 - G_1 \cdot \hat{s}_1 \quad , \quad y_2 = r_2 - G_2 \cdot \hat{s}_1$$

$$F(c) = \|y_1 - H_1 \cdot c\|^2 + \|y_2 - H_2 \cdot c\|^2$$

$$\hat{c}_1 = \underset{c \in C}{\arg\min} \, (F(c)) \quad , \quad \Delta_{c,1} = F(c)$$

If $(\Delta_{c,o} + \Delta_{s,o}) < (\Delta_{c,1} + \Delta_{s,1})$

$$(\hat{c}, \hat{s}) = (\hat{c}_o, \hat{s}_o)$$

Else

$$(\hat{c}, \hat{s}) = (\hat{c}_1, \hat{s}_1)$$

}

5   13. The receiver of claim 9 where said decoder for decoding a channel code is a trellis decoder.

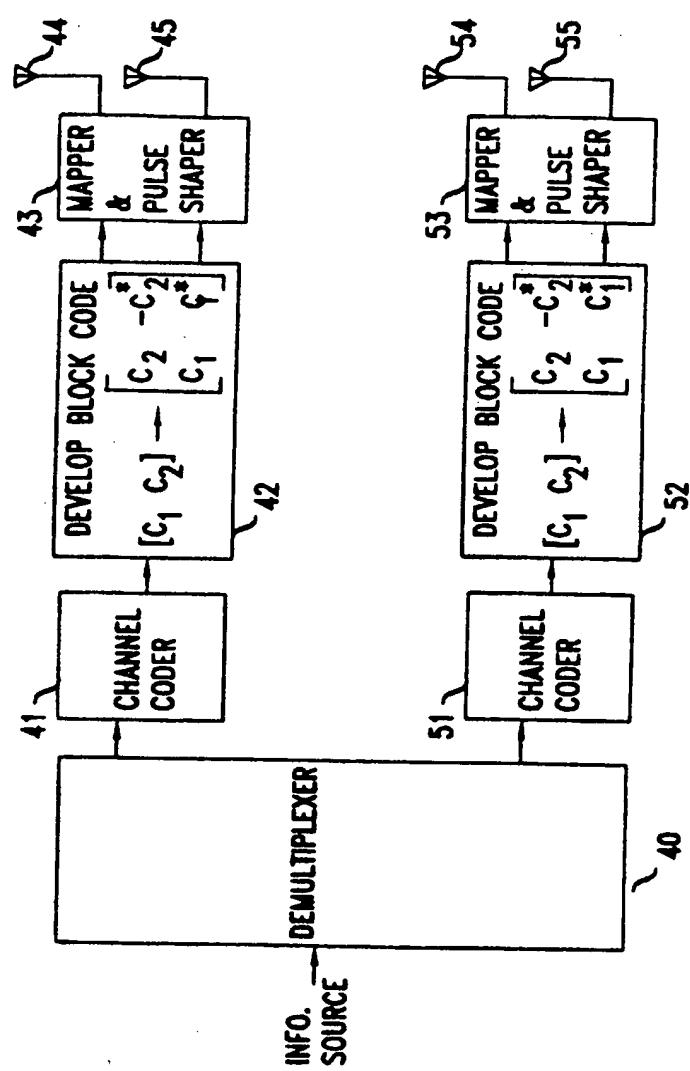14. The receiver of claim 9 where said decoder for decoding a channel code is a convolutional decoder.

10

FIG. 1

2/2

FIG. 2

INTERNATIONAL SEARCH REPORT

A. CLASSIFICATION OF SUBJECT MATTER

IPC 7   H04L1/00   H04L1/06

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

IPC 7   H04L

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category *	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	WO 97 41670 A (AT & T CORP) 6 November 1997 (1997-11-06) page 33, line 1 - line 21	1,2,9, 10,13,14
A	figure 21	3-8,11, 12
X	NAGUIB A F ET AL: "Space-time coded modulation for high data rate wireless communications" GLOBAL TELECOMMUNICATIONS CONFERENCE (GLOBECOM'97), PHOENIX, AZ, USA, 3 - 8 November 1997, pages 102-109, XP000737596 US, NEW YORK, IEEE ISBN: 0-7803-4199-6	1,2,9, 10,13,14
A	section 3.2 figure 5	3-8,11, 12

—/—

[X] Further documents are listed in the continuation of box C.	[X] Patent family members are listed in annex.

Date of the actual completion of the international search	Date of mailing of the international search report
13 December 1999	17/12/1999

Authorized officer

Ghigliotti, L

2

INTERNATIONAL SEARCH REPORT

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication,where appropriate, of the relevant passages	Relevant to claim No.
A	TAROKH V ET AL: "Space-time codes for high data rate wireless communication: performance criterion and code construction" IEEE TRANSACTIONS ON INFORMATION THEORY,US,IEEE INC. NEW YORK, vol. 44, no. 2, March 1998 (1998-03), pages 744-765, XP002089112 ISSN: 0018-9448 cited in the application the whole document	1-14
X,P	WO 99 23766 A (AT & T WIRELESS SERVICES INC) 14 May 1999 (1999-05-14) page 4, line 25 - line 30 claims 1-5	1,2,9, 10,13,14

2

INTERNATIONAL SEARCH REPORT

Information on patent family members

Patent document cited in search report		Publication date	Patent family member(s)		Publication date
WO 9741670	A	06-11-1997	AU	2744097 A	19-11-1997
			CA	2252664 A	06-11-1997
			EP	0906669 A	07-04-1999
WO 9923766	A	14-05-1999	EP	0960487 A	01-12-1999